A New QEA Based High Performance Sensorless Control of IM Drive

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***Abstract*— Quantum Evolutionary Algorithm (QEA) based proportional-integral (PI) controller tuning is used for getting fast speed response induction motor (IM) drive. A high performance simple sensorless control of induction motor drive is also presented in this paper. The control principle is based on direct torque control (DTC) with space vector modulation (SVM) technique. The SVM reduces the torque and flux ripples and improve steady state performance. A correlated real time recurrent learning (CRTRL) algorithm based on recurrent neural network (RNN) is used to estimate rotor flux which eliminates rotor position sensor. A simple speed estimator for the induction motor is proposed to make the controller cost effective and speed sensorless. The proposed control methodologies and simulation results are given and discussed.**

***Index Terms*— Sensorless, Quantum Evolutionary Algorithm, fast speed response, torque ripple, flux ripple.**

I. Introduction

High Performance Control (HPC) of induction motors is an interesting area for research and has wide applications in industries. The main objective of high performance controller is to obtain fast dynamic response of the drive system. It is so designed that it becomes less sensitive to motor and controller parameter perturbations and require minimum hardware for its practical implementation.

The field orientation control (FOC) is widely used in high performance motion control of induction motors. Because of torque and flux decoupling, FOC achieved good dynamic response and accurate motion control as separately excited dc motors. However, FOC has several disadvantages such as massive computational requirement, highly parameter and speed dependency. Accurate co-ordinate transformations [1-2] is an integral part of field-orientation control. Also, the DTC technique has gained wide acceptance in motor drive. But, it suffers notable torque, flux and current pulsations during steady state operation [3]. To achieve constant switching frequency operation, [4] incorporates a “torque ripple minimization” controller into the basic DTC control structure. Another method to produce DTC-SVM is proposed by Lascu et al. [5], wherein the flux controller produces a d-component of voltage and the torque controller a q-component. PI controllers are used in both the cases. To exploit the benefits of sensorless control, the speed estimation methods must achieve robustness against model and parameter uncertainties. Parameters of particular concern in the sensorless control literature are frequency-dependent R_r and temperature-dependent R_s and the load torque, all of which are very effective on the accurate estimation of flux and speed. To address the parameter sensitivity problem in induction motor speed sensorless control, a variety of approaches have been proposed [6]-[8]. But low speed estimation is a problem there. A lyapunov-function based flux and speed observer was developed [9] which can estimate R_s but not R_r. Duran et al. [10] performed a thermal-state estimation to compensate for the parameter and hence speed deviations due to heating. All researchers mentioned above proposed sensorless controller for induction motor drive and they didn’t consider the saturation effects in induction motor drive. But it is well known that consideration of saturation effect improves the dynamic characteristics of induction motor drive.

QEA is a novel probability optimization algorithm based on the concept and principles of quantum computing [11]. Compared with conventional evolutionary algorithm, QEA has a better characteristic of diversity in the population and can keep the balance of exploration and exploitation more easily- even with a small population. So QEA has become a research hotpot in recent years for optimizing the data.

In this paper, a new DTC-SVM based high performance sensorless control of induction motor drive is presented. To overcome the torque and flux ripple problem in DTC, SVM based inverter is used in which two switching sequences are used depending on the value of the reference voltage vectors. A simple and robust speed observer is developed to design the speed sensorless control of the induction motor drive. A CRTRL algorithm [12-13] based flux estimator is used to make the drive position sensorless. Finally, PI controller is tuned using QEA for fast speed response induction motor drive.

II. Mathematical Model

The dynamic model of induction motor can be represented in the synchronous reference frame (d-q) as:

$$v_{ds} = (R_s + pL_s)i_{ds} - L_s\omega_e i_{qs} + pL_m i_{dr} - L_m\omega_e i_{qr} \qquad (1)$$

$$v_{qs} = L_s\omega_e i_{ds} + (R_s + pL_s)i_{qs} + L_m\omega_e i_{dr} + pL_m i_{qr} \qquad (2)$$

$$0 = pL_m i_{ds} - L_m\omega_{sl} i_{qs} + (R_r + pL_r)i_{dr} - L_r\omega_{sl} i_{qr} \qquad (3)$$

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$$0 = L_m\omega_{sl}i_{ds} + pL_m i_{qs} + L_r\omega_{sl}i_{dr} + (R_r + pL_r)i_{qr} \tag{4}$$

$$T_e = Jp\omega_m + B\omega_m + T_L \tag{5}$$

Where, the symbols have their usual meanings.

L_m is used as a variable magnetizing inductance to consider the saturation effect in induction motor and shown in the appendix.

The developed electromagnetic torque in terms of *d*- and *q*- axes components is given by:

$$T_e = \frac{3}{2}P_p L_m\left(i_{qs}i_{dr} - i_{ds}i_{qr}\right) \tag{6}$$

Where P_p is the number of pole pairs.

Components of rotor flux are:

$$\psi_{dr} = L_r i_{dr} + L_m i_{ds} \tag{7}$$

$$\psi_{qr} = L_r i_{qr} + L_m i_{qs} \tag{8}$$

From (7) and (8), *d*- and *q*- axes rotor currents are:

$$i_{dr} = \frac{1}{L_r}(\psi_{dr} - L_m i_{ds}) \tag{9}$$

$$i_{qr} = \frac{1}{L_r}(\psi_{qr} - L_m i_{qs}) \tag{10}$$

Substituting (7)-(10) into (3) and (4) yields:

$$\frac{d\psi_{dr}}{dt} + \frac{R_r}{L_r}\psi_{dr} - \frac{L_m}{L_r}R_r i_{ds} - \omega_{sl}\psi_{qr} = 0 \tag{11}$$

$$\frac{d\psi_{qr}}{dt} + \frac{R_r}{L_r}\psi_{qr} - \frac{L_m}{L_r}R_r i_{qs} + \omega_{sl}\psi_{dr} = 0 \tag{12}$$

If the field orientation is established such that *q*-axis rotor flux is set zero, and *d*-axis rotor flux is maintained constant then equations (11), (12), (9), (10) and (6) becomes:

$$\psi_{dr} = L_m i_{ds} \tag{13}$$

$$\omega_{sl} = \frac{1}{\tau_r}\frac{i_{qs}}{i_{ds}} \tag{14}$$

$$i_{dr} = 0\ ,\ i_{qr} = -\frac{L_m}{L_r}i_{qs} \tag{15}$$

$$T_e = \frac{3}{2}P_p\frac{L_m}{L_r}\psi_{dr}i_{qs} \tag{16}$$

Where $\tau_r (= L_r / R_r)$ is the time constant of the rotor. Hence, only *q*-axis stator current controls the developed electromagnetic torque.

Substituting (13) and rearrangement of (1) and (2), yields:

$$v_{ds} = R_s i_{ds} + \frac{L_s}{L_m}\frac{d\psi_{dr}}{dt} + L_m\sigma\frac{di_{dr}}{dt} - \omega_e\frac{L_m}{L_r}\psi_{qr} - \omega_e\sigma L_s i_{qs} \tag{17}$$

$$v_{qs} = R_s i_{qs} + L_s\sigma\frac{di_{qs}}{dt} + \omega_e L_s i_{ds} + \frac{L_m^2}{L_r}\frac{di_{qs}}{dt} + \omega_e L_m i_{dr} + L_m\frac{di_{qr}}{dt} \tag{18}$$

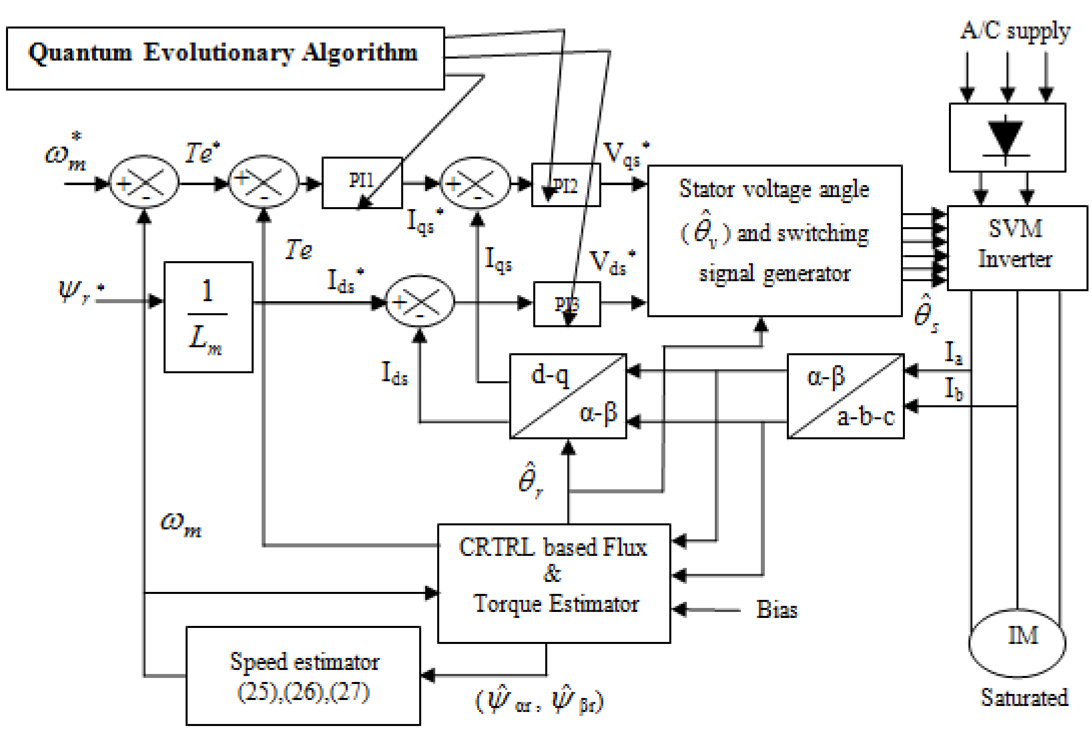

Figure 1. Proposed control scheme

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Where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ and $\omega_e = P_p \omega_m + \omega_{sl}$

III. PROPOSED CONTROL SCHEME

In the proposed scheme shown in Fig. 1, the reference torque T_e^* is generated from speed error maintaining minimum time and minimum loss control algorithm [14] to get the fast speed response. The two errors in i_{ds} and i_{qs} are processed through PI controllers to generate reference voltages v_{ds}^* and v_{qs}^*. The PI controllers are tuned using QEA to get further fast speed response. The stator command currents and command voltages are generated as follows:

$$i_{ds}^* = \Psi_r^* / L_m \qquad (19)$$

$$i_{qs}^* = k_{p1}\tau + k_{i1}\int \tau dt \qquad (20)$$

$$v_{qs}^* = k_{p2}(i_{qs}^* - i_{qs}) + k_{i2}\int (i_{qs}^* - i_{qs})dt \qquad (21)$$

$$v_{ds}^* = k_{p3}(i_{ds}^* - i_{ds}) + k_{i3}\int (i_{ds}^* - i_{ds})dt \qquad (22)$$

Here, τ is the torque error.

The angle of the input stator voltage needed with respect to the reference stator α -axis as:

$$\hat{\theta}_s = \hat{\theta}_v + \hat{\theta}_r \quad \text{Where, } \hat{\theta}_v = \tan^{-1}(v_{qs}^* / v_{ds}^*).$$

The magnetizing inductance representing the saturation in the magnetic circuit is given in the appendix and was found from the laboratory test of the motor.

IV. THE SPEED ESTIMATOR

The controller proposed in this paper needs the information of motor speed. First, the synchronous frequency can be found by noticing the angle of the rotor flux as:

$$\theta_r = \tan^{-1}\left(\frac{\psi_{\beta r}}{\psi_{\alpha r}}\right) \qquad (23)$$

Taking its derivative:

$$\dot{\theta}_r = \omega_e = \frac{\psi_{\alpha r}\dot{\psi}_{\beta r} - \psi_{\beta r}\dot{\psi}_{\alpha r}}{\psi_{\alpha r}^2 + \psi_{\beta r}^2} \qquad (24)$$

The rotor speed:

$$\omega_r = \omega_e - \omega_{sl} \qquad (25)$$

The rotor slip speed ω_{sl} can be expressed as follows:

$$\omega_{sl} = R_r\left(\frac{\psi_{\alpha r} i_{\beta s} - \psi_{\beta r} i_{\alpha s}}{\psi_{\alpha r}^2 + \psi_{\beta r}^2}\right) \qquad (26)$$

Substituting (24) and (26) in (25) the rotor speed can be estimated as:

$$\omega_r = \frac{1}{\psi_r^2}[\psi_{\alpha r}\dot{\psi}_{\beta r} - \psi_{\beta r}\dot{\psi}_{\alpha r} - R_r(\psi_{\alpha r} i_{\beta s} - \psi_{\beta r} i_{\alpha s})] \qquad (27)$$

V. QUANTUM EVOLUTIONARY ALGORITHM

The QEA is a stochastic search and optimization method based on the principles of natural biological evolution such as the quantum bit and the superposition of states [11]. The QEA can treat the balance between exploration and exploitation more easily when compared with CGA. The evolution procedure of QEA is shown in [11] clearly. Producing initial populations is the first step of QEA. The population is composed of the chromosomes that are represented by quantum bit or Q-bit which is the smallest unit of information in QEA. The corresponding evaluation of a population is called the "fitness function". It is the performance index of a population. In this paper, the fitness function is defined as:

$$Fitness = \frac{1}{\varsigma + 1} \qquad (28)$$

Where, $\varsigma = \int |speed\_error|dt$ is the objective function and $speed\_error = \omega_m^* - \omega_m$.

A Q-bit may be in the "1" state, in the "0" state, or in any superposition of the two. Superposition of logical state can be expressed as "$\alpha|0\rangle + \beta|1\rangle$". Another way of writing "superposition" as a vector is shown below:

$$\alpha|0\rangle + \beta|1\rangle \leftrightarrow \begin{pmatrix}\alpha \\ \beta\end{pmatrix} \qquad (29)$$

The complex numbers α and β are called the "amplitudes" of the superposition. $|\alpha_i|^2$ gives the probability that the Q-bit will be found in "0" state and $|\beta_i|^2$ gives the probability that the Q-bit will be found in "1" and they satisfy the normalization condition $|\alpha_i|^2 + |\beta_i|^2 = 1$. A Q-bit is also defined with a pair of numbers (α, β) and a Q-bit individual as a string of m Q-bits is defined as [11]:

$$q = \begin{bmatrix}\alpha_1 & \alpha_2 & \dots & \alpha_m \\ \beta_1 & \beta_2 & \dots & \beta_m\end{bmatrix} \qquad (30)$$

Where $|\alpha_i|^2 + |\beta_i|^2 = 1$, $i = 1, 2, 3\dots..m$. A Q-gate defined as a mutation operator is applied on the Q-bit to update their probability amplitudes as follows:

$$\begin{pmatrix}\alpha_i' \\ \beta_i'\end{pmatrix} = \begin{bmatrix}\cos(\Delta\theta_i) & -\sin(\Delta\theta_i) \\ \sin(\Delta\theta_i) & \cos(\Delta\theta_i)\end{bmatrix}\begin{pmatrix}\alpha_i \\ \beta_i\end{pmatrix} \qquad (31)$$

Where $\Delta\theta_i$, i=1, 2, 3…., m, is the rotation angle of a Q-bit towards the "0" state or "1" state depending on its sign. $|\alpha_i'|$ and $|\beta_i'|$ must satisfy the normalization condition $|\alpha_i'|^2 + |\beta_i'|^2 = 1$. The crossover operator is employed after a given interval of generations. Along with Q-bit population, a binary population is also maintained for evaluation process. In this paper, QEA is used to optimize the value of the gain coefficients of PI controllers.

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VI. SIMULATION RESULTS

The induction motor used for the simulation study has the following parameters: 0.75KW-50Hz-415V, stator resistance, R_s=13.25Ω, rotor resistance, R_r=16.818Ω, leakage inductance of the windings, L_s=L_r=0.7359H and mutual inductance, L_m=0.7114H moment of inertia, J=0.0075Kg-m^2 and damping coefficient, B=0.00107Nm-sec/rad.

The induction motor model described above and the control system described in Fig. 1 along with the estimation method was implemented in a pc based C++ environment. The flux estimator was trained off-line so that it estimates the flux components accurately.

A. Rotor Flux and Position Estimation

Fig. 2 shows the actual and estimated α-axis and β-axis rotor fluxes under both in transient and steady state conditions. It is apparent that the estimated responses are matching accurately with the actual responses. Thus the acceptability of the proposed CRTRL has been confirmed.

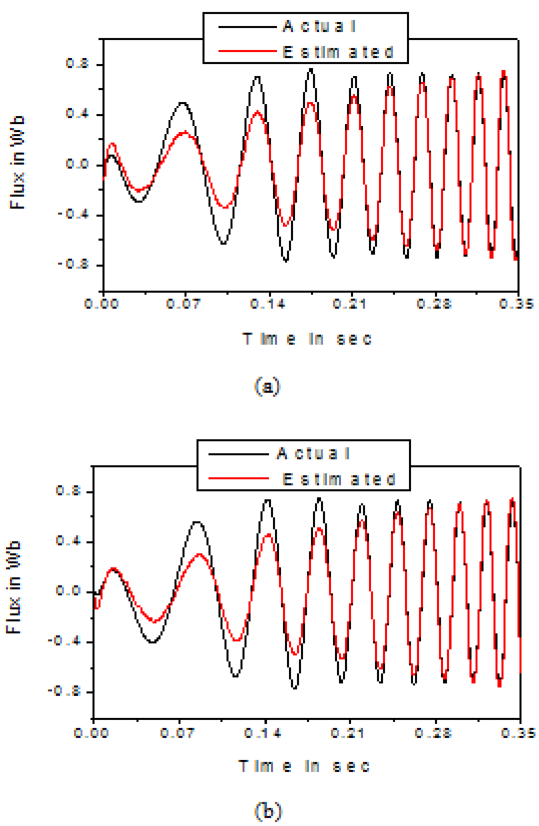

Figure 2. Actual and estimated responses of (a) α-axis rotor flux, and (b) β-axis rotor flux under both in transient and steady state conditions

B. Speed Estimation and Dynamic Behavior

Fig. 3 shows the estimated speed, torque and flux responses under transient and steady-state conditions with load torque 0.5 N-m. It is found that the proposed control scheme is capable to estimate the speed accurately at both in transient and steady-state conditions and shown by small scale representation in Fig. 3(a). It is confirmed that the proposed speed estimator is capable to estimate the speed even at very low speed accurately. From Figs 3(b) and 3(C), it can be seen that the proposed control system generates negligible torque and flux ripple due to the consideration of main flux saturation effect. Effectiveness of the proposed controller is tested by different set speed and shown in Fig. 4. Negligible overshoot and undershoot is present in the speed estimation and shown by small scale representation in Fig. 4. One important matter is noticeable here that if there is no change of speed, there is no overshoot and undershoot in the estimated speed.

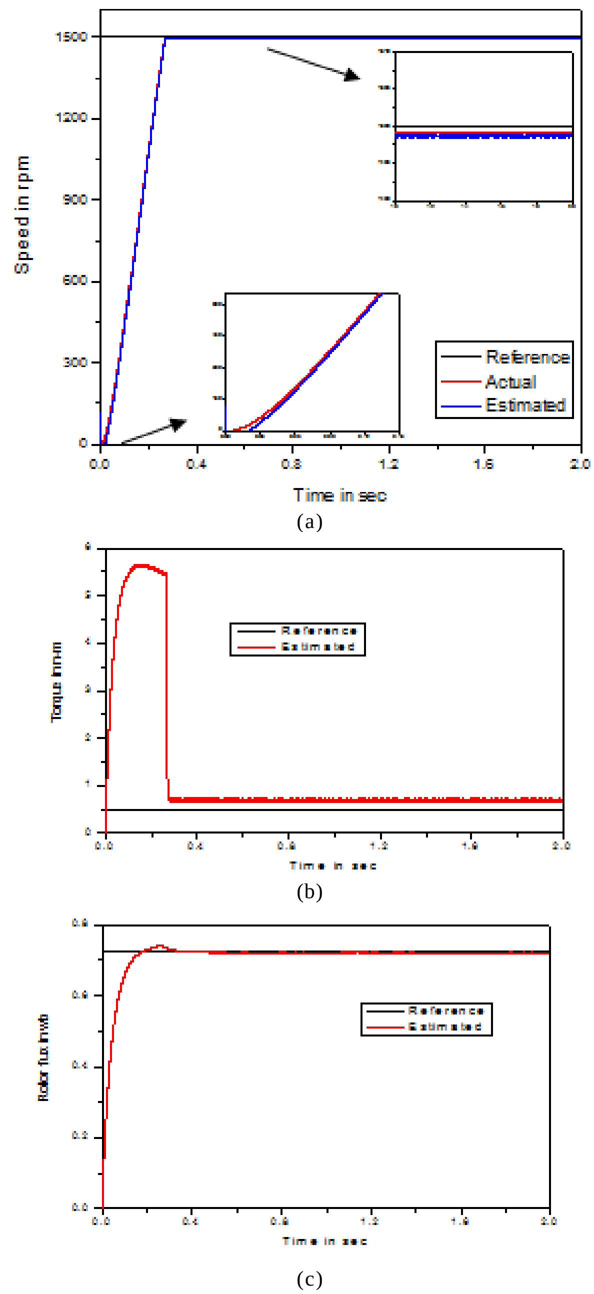

Figure 3. Simulated (a) actual and estimated speed response (b) estimated electromagnetic torque, and (c) estimated rotor flux under transient and steady-state conditions.

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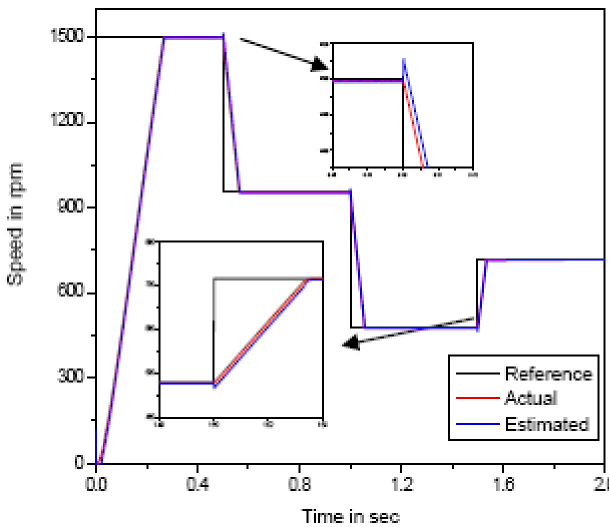

Figure 4. Simulated actual and estimated speed response under transient and steady state conditions with different set speed.

C. Fast Speed Response IM Drive

Fast speed response of induction motor drive is further achieved using QEA based PI controller tuning compared with conventional evolutionary algorithm based PI controller tuning and shown in Fig. 5. Effectiveness of the controller is also tested by different set speed.

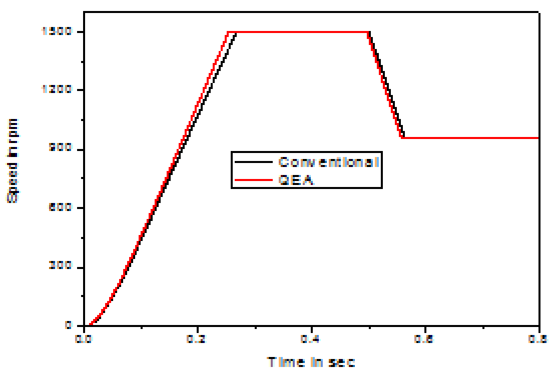

Figure 5. Speed response of IM drive using conventional PI controller and QEA based PI controller with different set speed

Conclusion

This paper has provided a new DTC-SVM based high performance sensorless control method for induction motor drive. CRTRL algorithm based rotor flux estimator eliminates the rotor position sensor. A simple speed estimator is used to make the proposed controller speed sensorless, more cost effective and practical. Consequently the flux, torque and speed estimation is accurate and hence motor performance is improved. A minimum time minimum loss algorithm realizes fast speed response IM drive. Further fast speed response IM drive is achieved using QEA based PI controller tuning.

APPENDIX

Saturation characteristic for induction machine:

Mutual inductance, $L_m = f(i_{mag})$

$$= b_0 + b_1 i_{mag} + b_2 i_{mag}^2 + b_3 i_{mag}^3 + b_4 i_{mag}^4 + b_5 i_{mag}^5$$

Here, b_0 =0.9328; b_1 = 0.27058; b_2 =-0.95898; b_3 =0.54123; b_4 =-0.5479; and b_5 =-0.01859.

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